Frequency Modulation of a Millimeter-wave IMPATT Diode Oscillator and Related Harmonic Generation Effects

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In this paper we report the performance of a continuous wave millimeter-wave IMPATT diode oscillator with a wide-band tunability. The diode is mounted in an iris wafer circuit; its oscillation frequency can be modulated either by a varactor diode or by direct modulation of the IMPATT diode bias current. The oscillator has been successfully used as a millimeter-wave frequency deviator in an experimental pulse code modulation millimeter-wave system.

We also report detailed measurements on subharmonic frequencies in IMPATT diode oscillators. Experimental results show that wide frequency tunability can be obtained with a circuit which provides an "idler" resonance at one-half the fundamental transit-time frequency. The results also show that by providing "idler" resonances at both the transit-time frequency and at one-half of the transit-time, the oscillation at $\frac{3}{2}$ the transit-time frequency is enhanced and yields a useful output power of 2 mW at 86 GHz.

I. INTRODUCTION

A continuous wave (CW) millimeter-wave silicon IMPATT diode oscillator was used satisfactorily as a local oscillator in an experimental pulse code modulation (PCM) millimeter-wave repeater system.[1, 2] That oscillator circuit used a radial-line resonant cavity whose resonant frequency was the primary factor determining the oscillation frequency. The oscillator was difficult to tune either mechanically or electronically.

While such characteristics are desirable for fixed frequency local oscillator applications, other applications which demand wideband

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performance and tunability are also of interest. For example, a millimeter-wave frequency deviator using an IMPATT diode would be advantageous over the L-band deviator and up-converter combination in the PCM repeater system. The IMPATT diode oscillator could deliver power at least one order of magnitude greater than the up-converter.

A previous paper showed that the circuit inductance of the radial-line cavity in shunt with the diode was much smaller than the diode inductance.[1] Therefore, the oscillation frequency could be tuned only a few hundred megahertz by varying the diode bias current.* In addition, the radial-line cavity was loosely coupled to the external waveguide circuit so that a tuning range of only 300 MHz was obtained when a varactor diode was used for frequency deviation. To improve the tunability, it is necessary to have (*i*) the IMPATT diode equivalent inductance dominate that contributed by other circuit elements, and (*ii*) the diode closely coupled to controllable external circuit parameters.

It is important to notice that the tunability can be further improved by providing more than one resonant circuit for the oscillator. The frequencies are harmonically related, and oscillation at each of these frequencies is reactively terminated except the one for the output. Take as an example the two-frequency case. The tunability of the oscillator near the transit-time frequency (say 60 GHz) can be improved if a lossless resonant circuit at half the output frequency (30 GHz) is incorporated in the oscillator circuit. The oscillation at 30 GHz is terminated by a cut-off waveguide and therefore is designated as the "idler," an analog to the idler in harmonic generators using varactor diodes. It is not necessary that the low frequency be used as the idler. Swan has shown that by providing an idler resonance at twice the transit-time frequency, improvements in both the power output and the tunability can be obtained.[3]

To achieve wide band tunability, IMPATT oscillators were designed using resonant-iris structures. It can be shown that the diode equivalent inductance is dominating in this circuit compared with that of the radial-line cavity structure. The loaded Q of the iris is about 10 which provides a wide bandwidth for oscillation. The oscillation frequency could be tuned over 9 GHz in the 50 to 60 GHz range by varying the bias current of the IMPATT diode, thus varying the diode

* The diode equivalent inductance is inversely proportional to the diode bias current.

equivalent inductance. It could also be tuned over 3 GHz by using a varactor diode closely coupled to the IMPATT diode.

Notice that by using the fundamental transit-time frequency as the idler, the power output at one-half the transit-time frequency is, in general, higher than the power output when the half frequency is used as the idler in our particular circuit. We refer to the frequency which is one-half the transit-time frequency as the subharmonic frequency throughout this paper.

In the sections which follow we describe the circuit structure of the oscillator, the performance as a tunable oscillator, and the results of frequency modulation of the oscillator in an experimental millimeter-wave PCM repeater system. Then measurements of various harmonic frequencies existing in the oscillator circuit, and the identification and effect of the subharmonic oscillation are described in detail.

II. DESCRIPTION OF THE OSCILLATOR

Figure 1a shows the resonant-iris structure. The iris is made in wafer form similar to the Sharpless wafer.[4] The range of the oscillation frequency and the Q of the resonant iris are determined by the size, thickness, and shape of the iris aperture. Refs. 5, 6, and 7 give details of iris characteristics. The wafers used were 0.100 inch thick,

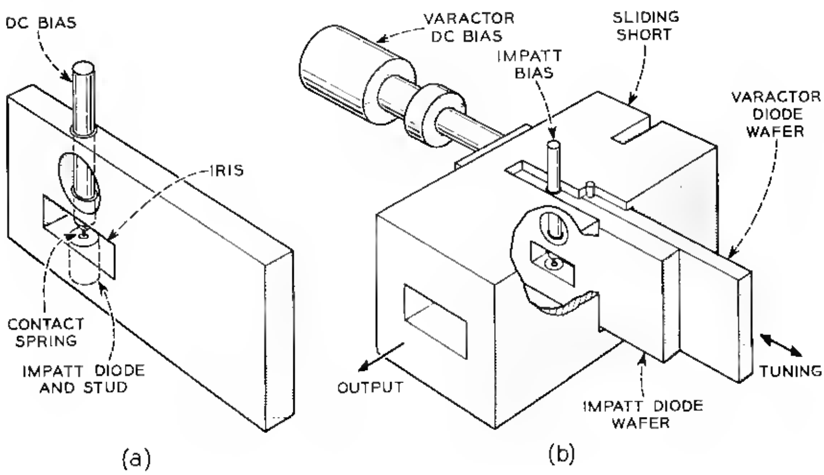

Fig. 1—Artist's view of the oscillator assembly. (a) IMPATT diode wafer. (b) Oscillator mount.

and had a rectangular aperture 0.010 or 0.030 inch high and 0.100 to 0.148 inch wide.

The IMPATT diodes had a 0.001 inch diameter mesa structure and nearly abrupt junctions as reported by Misawa.[8] Typical diode characteristics are given in Table I. The diodes were thermal-compression bonded onto gold-plated copper studs 0.063 inch in diameter. After bonding, the diodes were coated with a thin layer of silicone varnish which enchances the mechanical strength of the mesa structure. Electrical contact to the diode was made by a 0.032-inch nickel rod with a welded contact spring, as pictured in Fig. 1a. For the varactor-tuned oscillator, a varactor diode in a Sharpless wafer was mounted adjacent to the iris wafer as shown in Fig. 1b. The relative coupling between the IMPATT diode and the varactor diode (thus the tuning characteristic) was adjusted by varying the relative position of the two diodes in the plane of the cross section of the waveguide.[9] The complete assembly in a RG-98/U waveguide is shown in Fig. 1b.

Table I — Typical Diode Characteristics*

Epitaxial layer thickness†	2.1 μ
Epitaxial layer doping density‡	6×10^{16} cm^{-3}
Junction depth†	1.0 μ
Space charge layer width‡	0.75 μ
Breakdown voltage (at 1 μA)	19V
Capacitance at breakdown	0.19 pF
Junction area A	1.5×10^{-5} cm^2

* All diodes were from the LO 1114 series.
† Measured by staining and interference fringe.
‡ Obtained from voltage dependence of capacitance.

III. OSCILLATOR PERFORMANCE

Figure 2 shows the typical performance of an oscillator without the varactor. (The varactor wafer is replaced by a blank wafer.) The iris used had an aperture of 0.130 by 0.030 inch. The CW ouput power (above 1 mW) and the frequency are plotted as a function of the IMPATT diode current. Oscillation began at a diode current of about 50 mA compared with 100 mA for the cavity structure.[1] The output power was optimized at each current level by adjusting an E-H tuner in front of the diode and a sliding short in back of each

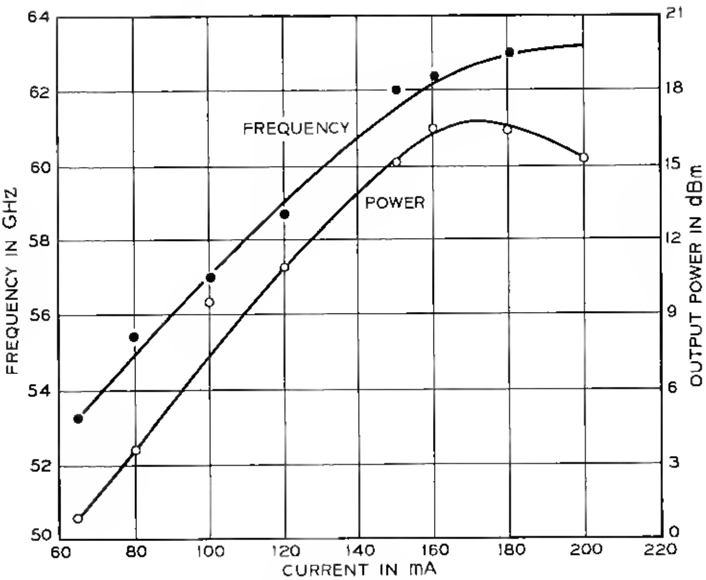

Fig. 2 — Maximum output power and frequency as a function of diode current.

diode. Note that the tunable band (about 10 GHz) is considerably wider than that previously reported for a radial line cavity structure. By varying the bias current with fixed mechanical tuning, the frequency could be tuned over 4 GHz with a 3 dB variation in the output power, and over 9 GHz with a minimum output power of 1 mW.

Table II summarizes the results of various circuits with different iris apertures. Although there are not enough data to draw a meaningful conclusion, it seems that the oscillation frequency was relatively independent of the iris width compared with iris height, which implies that the inductance of the center post (with contacting spring) partially controls the oscillation frequency. This statement is based on the assumption that diodes used in the test are of uniform characteristics since they were made from a single slice and batch processed. The dc characteristics of the diodes showed less than 5 percent variation in capacitance and in breakdown voltage.

IV. FREQUENCY MODULATION

As mentioned in Section I, one goal in this work was to design an IMPATT diode oscillator circuit which could be frequency modulated

TABLE II—RESULTS OF CIRCUITS WITH VARIOUS DIODES AND IRIS APERTURES

Diode No.	Width (in)	Height (in)	C_B‡ (pF)	Tunable¶ bandwidth (GHz)	P_{max} (dBm)
1	148	0.030	0.19	50 – 60	15
2	130	0.030	0.19	53 – 63	16.5
3	120	0.030	0.19	52 – 59	11.5
4	110	0.030	0.18	56 – 63.2	9.4
5	100	0.030	0.19	53 – 61	10.7
6	130	0.010	0.16	64 – 70	11.5
7	148	0.010	0.18	58 – 65	11.1

‡ C_B is the diode capacitance measured at breakdown ($V_B = -19$ volts).
¶ The lower limit on frequency is arbitrarily chosen as the frequency for which the power is 1 mW. The upper frequency limit occurs at a bias current of 160 mA which is well below the burn-out level for the LO 1114 series diodes.

and used as a frequency deviator in a PCM millimeter-wave repeater system.[2] Two approaches were taken. The first used a varactor diode to tune the circuit susceptance outside the diode; the second used the fact that the diode inductance (from avalanching) varies inversely with the bias current.

Both methods have advantages and disadvantages. The power output of the varactor-tuned oscillator remains almost constant over the frequency band so that amplitude modulation is negligible. However, to achieve ultimate performance the circuit is much more complex. The bias-current tuned oscillator has simpler circuitry, but inevitably its power output varies with bias current, which results in AM distortion. However, the AM distortion can be overcome by proper tuning of the circuit if the modulation index is small. (See Section 4.2)

4.1 *The Varactor-Tuned FM Deviator*

The varactor tuned oscillator is shown in Fig. 1b. The varactor diodes used were planar diffused GaAs diodes with a honeycomb structure.[10] The zero-bias capacitance was 0.04 – 0.05 pF and the breakdown voltage was 20 volts. The capacitance varied with voltage approximately as $C = C_0\ (1 + V/\varphi)^{-0.4}$, where C_0 is the junction capacitance at zero bias, V is the bias voltage, and φ is "built-in" voltage which is approximately 1 volt for GaAs. The diode was then mounted in a Sharpless wafer which in turn was inserted into the oscillator mount. The varactor diode was about 0.080 inch behind the IMPATT diode. The coupling between the varactor diode and the

IMPATT diode was tuned by sliding the varactor wafer relative to the IMPATT wafer. Additional tuning was provided by a sliding short behind the varactor diode and an E-H tuner in front of the IMPATT diode.

The self-resonant frequency of the varactor diode, as measured by the transmission resonance technique, was used to correlate with the tuning sensitivity of the oscillator. The self-resonant frequency can be varied by changing the length of the contacting wire in the varactor wafer.[11] It was found experimentally that the best tuning sensitivity was obtained when the varactor diode was self-resonant near the "idler" frequency.

A typical frequency tuning characteristic of the oscillator is sbown in Fig. 3 where the frequency is plotted as a function of tbe dc bias voltage (reverse biased) on the varactor. A tunable band of 2.5 GHz was obtained. Notice that the linear region was 1.5 GHz witb a sensitivity of 1 GHz per volt. Also notice that the power output varied ±1 dB over the tunable band.

Frequency modulation with sinusoidal drive on the varactor was achieved at both the 160 MHz and the 500 MHz modulation fre-

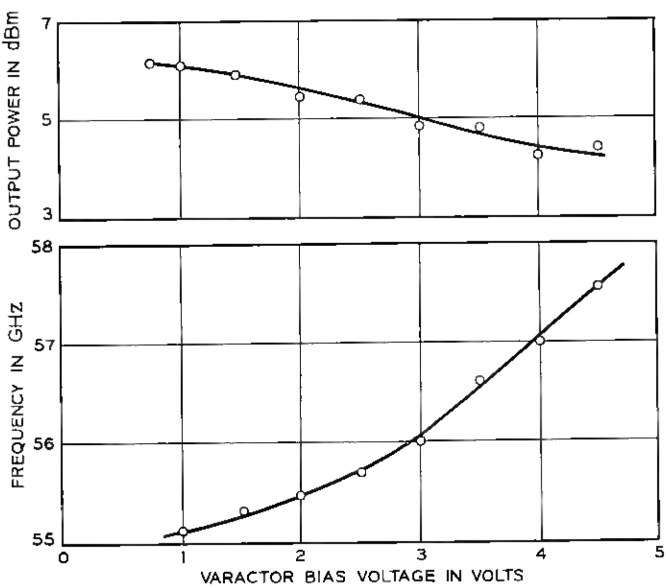

Fig. 3 — Performance of the varactor-tuned oscillator.

quency. Figure 4 shows a series of the FM spectrum at 160 MHz modulation as the modulating power is increased. For the particular diode tested, the minimum modulating power required at 160 MHz for complete carrier suppression was about 4 mW (to a 50 ohm load).

4.2 *The Bias-Current Tuned FM Deviator*

The second method of frequency modulation was achieved by direct modulation of the IMPATT diode bias current (the varactor diode wafer was replaced by a blank wafer). The modulating signal was directly applied to the bias through a coupling capacitor. Tbe dc tuning sensitivity was between 100 to 200 MHz per milliampere at optimum tuning conditions; this resulted in a decrease of 1 dB in output power compared with the maximum power obtained without frequency deviation. For a sinusoidal drive at 160 MHz, complete carrier suppression was achieved at 0.25 mW drive power. Figure 5 shows the spectrum of the modulated signal at various drive levels. When it is tuned properly, no appreciable AM distortion was seen at low modulation index.

4.3 *System Performance with the IMPATT Diode Oscillator-Deviator*

The methods of IMPATT diode oscillator frequency deviation were tested in a two-level PCM millimeter-wave repeater system.[2] The error rate of the system with the IMPATT diode oscillator used as the deviator was measured as a function of the signal-to-noise ratio. Figure 6 is a simplified block-diagram of the test circuit. The IMPATT diode oscillator-deviator is driven by a random-word generator either through the bias of the varactor for varactor-tuned deviation or through the bias of the IMPATT diode for current-tuned deviation.

The random word generator produces 160 megabit pulses per second with a random distribution in polarity. These pulses modulate either the IMPATT diode current or the varactor diode bias to deviate the millimeter-wave output frequency according to the input pulse polarity. The IMPATT diode oscillator-deviator output is down converted to an IF signal in the 1.3 ± 0.25 GHz band. The IF signal is then amplified and injected through a 500 MHz bandwidth filter to a phase-locked oscillator. The function of this oscillator is to act as a limiter thus removing amplitude modulation from the frequency modulated signal. The output of the phase-locked oscillator is further amplified before being injected into the differential phase detector and timing recovery circuit. The latter gives two output voltages. The

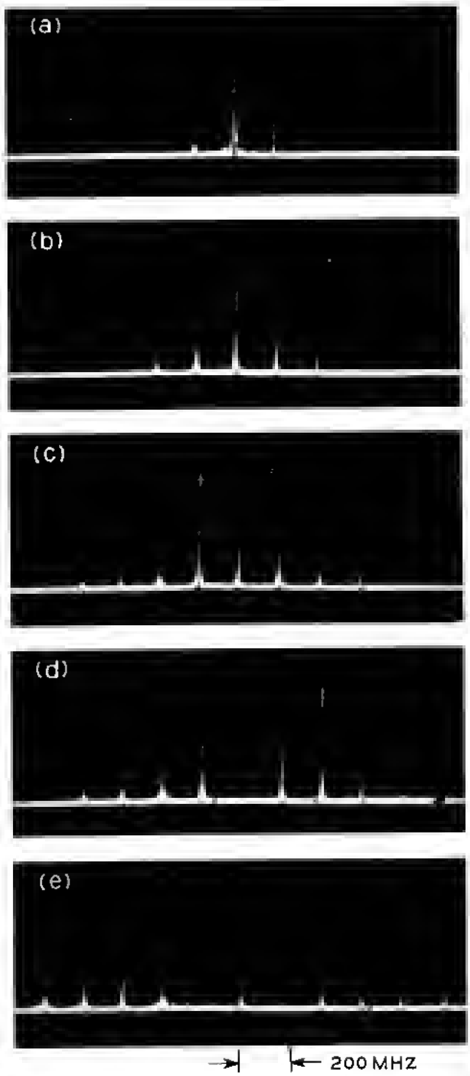

Fig. 4 — FM spectrum at 160 MHz modulation of varactor-tuned oscillator. Complete carrier suppression is shown in (d) with 6 dBm IF power. Others are (a) −9 dBm, (b) −4 dBm, (c) 1 dBm, and (e) 11 dBm (first sideband suppression).

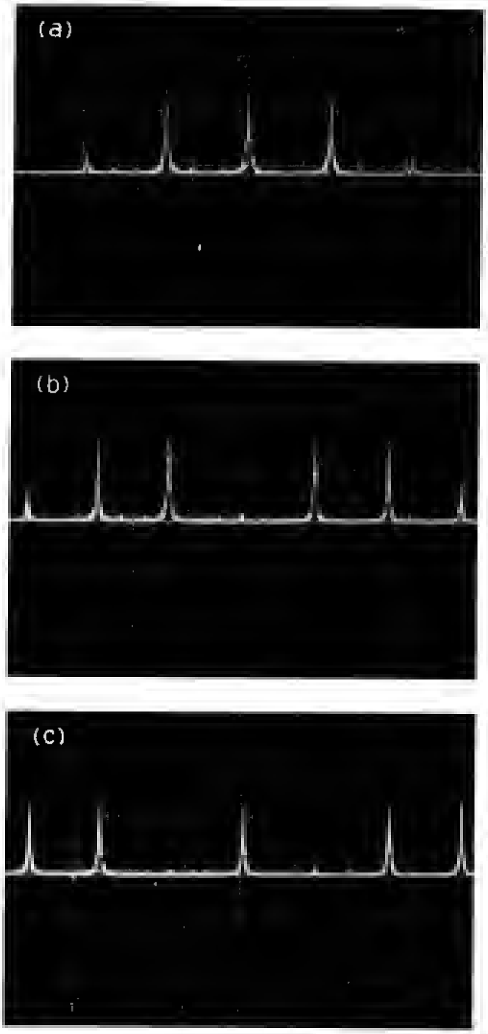

Fig. 5 — FM spectrum at 160 MHz modulation on the bias current. (a) IF power of −11 dBm for equal sidebands and carrier, $\Delta f/f = 1.435$, (b) IF power of −6 dBm for carrier suppression $\Delta f/f = 2.405$, and (c) IF power of −2 dBm for first sideband suppression, $\Delta f/f = 3.832$.

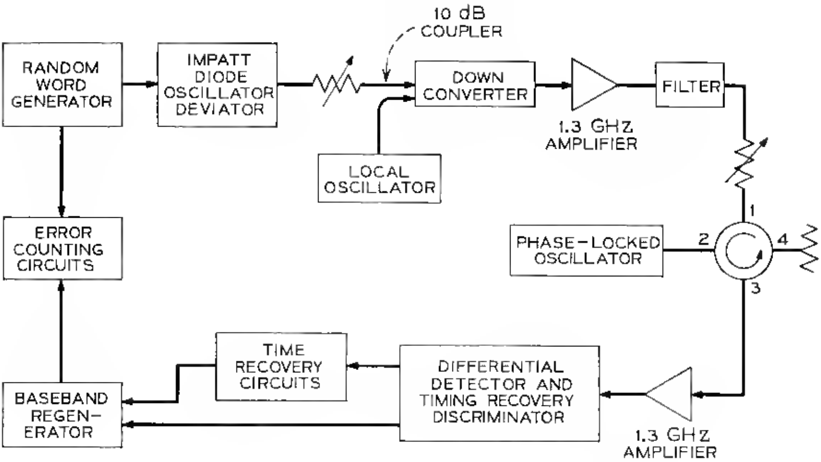

Fig. 6 — Circuit block diagram for the error rate measurement.

first output is ideally a replica of the baseband drive signal from the random word generator. The second output provides a timing signal for the regenerator, which samples the polarity of the first input at the time of arrival of the second input. The regenerator output is then compared with the random word generator output in an error eounting eireuit. (Referenee 12 gives further details of circuit eomponents.)

For the random-word modulation just deseribed, the drive required at the 160 megabit rate was about 0.1 volt peak-to-peak (to a 75Ω load) for the varaetor-tuned deviation seheme, and 0.03 volt peak-to-peak for the current-tuned deviation scheme at the optimum error-rate. Figure 7 eompares typical eye-diagrams for the two deviation methods tested. For the latter ease, the measured error-rate was 10^{-7} for a 16 dB signal-to-noise ratio, which was 2 dB worse than that obtained with a 1.3 GHz tunnel-diode deviator.[2] The major problem appeared to be the FM noise on the IMPATT oseillator output when it was optimally tuned for best deviation. The noise eould have been redueed by inereasing the eireuit Q and eonsequently inereasing the IF drive power.

Error rate improvement could have been obtained by matehing the input impedanee to the IMPATT bias eireuit over the bandwidth of the baseband signal. A narrow band matehing results in distortion of the baseband pulses.

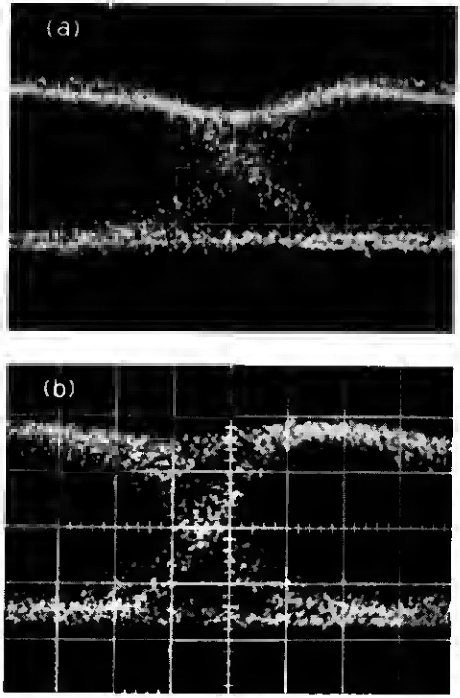

Fig. 7 — "Eye-diagrams" of the differentially-coherent-detected random word signal. (a) With the varactor-tuned FM deviator. (b) With the current-tuned FM deviator.

V. MEASUREMENTS OF SUBHARMONIC FREQUENCIES

As mentioned in Section I, the oscillator circuit under study provides at least two resonant circuits for the IMPATT diode, namely, an output circuit which is resonant at the transit-time frequency in M-band (50 to 75 GHz) and an "idler" circuit which is resonant at one-half the transit-time frequency in V-band (26.5 to 40 GHz). The two-frequency arrangement, with the frequencies harmonically related, has improved the tuning sensitivity of the oscillators, as we have shown in previous sections.

Although the existence of such a subharmonic oscillation can be deduced from the experiments described below, it is not surprising in view of the highly nonlinear nature of the impact ionization and avalanche process. Indeed, small-signal theories have predicted that the negative resistance exists over, at least, an octave in frequency.[13–15]

Parametric oscillations and mixing have been observed by DeLoach, Johnston, Evans, and Haddad.[16, 17] High-efficiency, low frequency, subtransit time oscillations have also been reported by Prager, Johnston, Scharfetter, and others,[18–19] and analyzed recently by Scharfetter and others.[20] In Scharfetter's analysis, the low-frequency oscillation is not necessarily harmonically related to the fundamental transit-time frequency. In recent work, Swan[3] showed that improvements in both output power and tuning bandwidth are obtained by providing an idler circuit which resonates at the second harmonic of the transit-time frequency. This result is similar to our findings except that in our case the idler is at one-half the transit-time frequency.

5.1 *Measurements*

Since the idler frequency is below the cutoff frequency of the RG-98/U waveguide, it is impossible to detect its existence directly at the output port of the oscillator as shown in Fig. 1. However, the existence of such a signal at the subharmonic frequency would result in mixing with the fundamental to produce an output at 3/2 the fundamental frequency. It was found that the next higher-order harmonic detected was always 3/2 of the oscillation frequency independent of bias current for all the diodes tested.

For direct detection of the below-cutoff subharmonic oscillation dielectric-filled waveguide tapers which had a cutoff frequency of 18 GHz were used as shown in Fig. 8a. However, a short section of air-filled RG-98/U waveguide with about 30 dB attenuation at 30 GHz remained between the IMPATT diode and the output waveguide taper. This arrangement retained the same subharmonic oscillation circuit conditions. Yet, if enough power exists for the signal at the subharmonic frequency to pass through the short section of air-filled RG-98/U waveguide, both the fundamental and the subharmonic frequencies should be present at the output port. A mixer and a spectrum analyzer were used to detect the subharmonic, while a wavemeter and a diode detector were used for the fundamental frequency. There was indeed an appreciable amount of power (estimated at 9 dBm at the diode) at the subharmonic of 27 GHz; the fundamental was exactly twice the subharmonic frequency, or 54 GHz, within measurement error.

When the short section of the RG-98/U waveguide preceding the diode wafer was also filled with dielectric the power output at the subharmonic was increased appreciably, but the frequency shifted

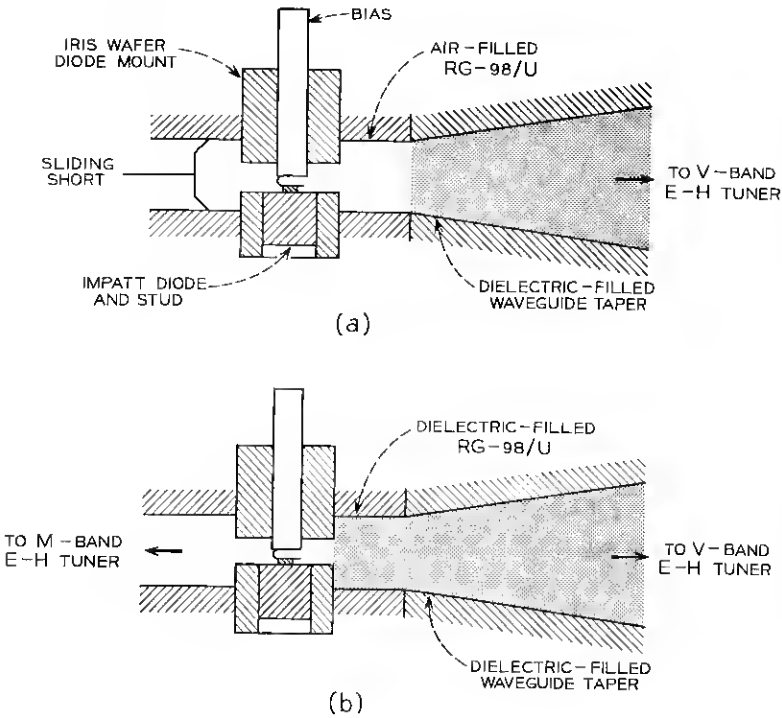

Fig. 8 — Waveguide arrangements for measuring the subharmonic signal.

slightly because of the change in loading impedance. The measurements of the subharmonic power on several diodes are summarized in Table III.

When the sliding short on one side of the oscillator mount was replaced by an E-H tuner followed by a wavemeter and a detector, as shown in Fig. 8b, both the fundamental and the subharmonic could be measured simultaneously. The results again confirmed that the subharmonic frequency was exactly one-half of the fundamental as shown in Fig. 9 for diode No. 3. This harmonic relation held true for any bias current.

The E-H tuner in the M-band circuit serves as an impedance matching device. By mismatching the oscillator to the load in the M-band circuit (thus reducing the 58 GHz power delivered to the load), the power delivered to the V-band load increases, and vice versa. Thus for efficient operation of IMPATT diodes, the subharmonic (and the harmonics)[3] should be reactively terminated. Likewise maximum sub-

TABLE III—MEASUREMENT OF POWER AT ONE-HALF THE TRANSIT-TIME FREQUENCY

Diode No.*	Bias current (mA)	Frequency (GHz)	Power output (dBm)
1	115	28.13	12.05
2	129	28.36	11.2
3	140	29.05	12.0
5	140	29.51	10.2
6	150	34.48	16.9
7	150	32.25	15.2
8	150	32.98	14.7
9	150	31.91	11.1

* The diode number is consistent with that in Table II.

harmonic power can be obtained when the fundamental and all harmonics are reactively terminated. Using this approach, by reactively terminating both the fundamental and the subharmonic, we obtain an output power at 3/2 the transit-time frequency of about 3.3 dBm at 86 GHz.

5.2 *Harmonic Phase-Locking*

The Fig. 8b circuit arrangement also was used for harmonic phase-locking. The locking-signal was injected in the V-band end through

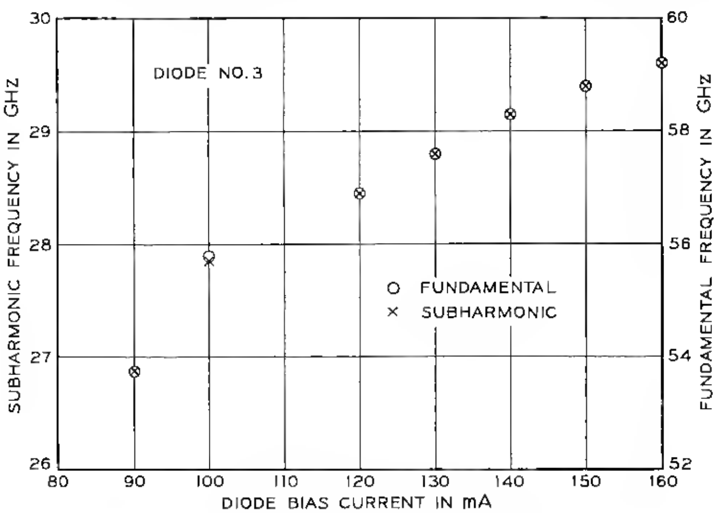

Fig. 9—Fundamental and subharmonic frequencies as a function of diode bias current.

a 10 dB directional coupler. Both the locked signals at the V-band and at the M-band were measured. The gain-bandwidth product $(2\Delta f/f_0)\ (P_0/P_i)^{1/2}$ was about 0.05 to 0.07 for the V-band output.* The M-band output frequency was simultaneously phase-locked witb the locked-bandwidth exactly twice that of the V-band. Since the power output in the M-band was about 1 mW compared with about 6 mW at V-band, the apparent gain bandwidth product was much less.

5.3 *V-Band Circuit with Cap Structure*

The experiments of Section 5.1 were conducted on IMPATT diodes in tbe iris circuit sbown in Fig. 1. The same behavior was observed in a different circuit structure. A V-band oscillator was constructed using a resonant cap structure similar to the circuit described in Ref. 1. (see Fig. 10.) Caps were made in various diameters, and could be

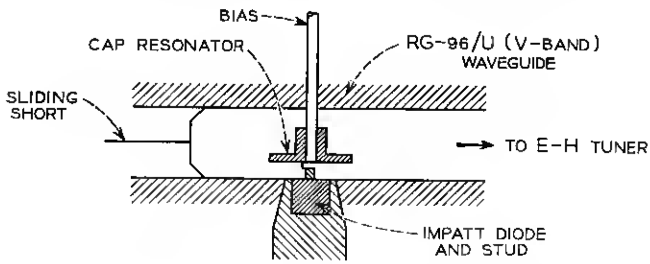

Fig. 10 — A V-band IMPATT oscillator using cap resonator structure.

slid up and down the center bias rod, thus permitting the frequency of oscillation to be varied. A diode with very similar characteristics (breakdown voltage and capacitance) as the ones used before was selected from tbe same batch (LO 1114). The fundamental oscillation was in the range of 64 to 72 GHz with an output power of 10 dBm at 67 GHz for 150 mA bias. The subharmonic power was also detected.

To optimize the fundamental power, a V-band to M-band taper was used in addition to an E-H tuner so that the subharmonic oscillation was reactively terminated. Table IV shows the results. Notice that when the diode was biased at the same current level, the subharmonic frequency was close to ½ the fundamental frequency. The

* Here we define: f_0 = free running oscillation frequency
P_0 = free running oscillation output power
P_i = injected power at frequency $f_0 \pm \Delta f$
$2\Delta f$ = bandwidth over which the oscillator is phase-locked.

TABLE IV—OSCILLATION IN V-BAND CIRCUIT WITH CAP-RESONATOR STRUCTURE

Subharmonic			Fundamental (Subharmonic reactively terminated)		
Bias (mA)	Frequency (GHz)	Power output (dBm)	Bias (mA)	Frequency (GHz)	Power output (dBm)
80	32.75	7.5	80	64.95	4.0
100	32.87	9.5	105	65.7	5.7
150	33.4	14.0	150	67.0	10.5
160	36.0	14.0	160	72.0	11.1

slight difference resulted from the reactive termination of the subharmonic which required slight retuning for maximum output power at the fundamental frequency.

5.4 *Comparison of the Outputs*

Table V compares the fundamental and subharmonic power outputs in the two different circuits. For the iris-wafer circuit the diode was mounted in the RG-98/U waveguide; dielectric-filled waveguide and tapers were used when the subharmonic frequency power was measured. The dielectric-filled waveguide and tapers had an insertion loss of 1 dB in the frequency range of interest, which was taken into account for the power listed in the fourth column in Table V. For the resonant cap circuit, the diode was mounted in the V-band wave-

TABLE V—COMPARISON OF FUNDAMENTAL AND SUBHARMONIC POWER OUTPUTS

Diode No.	Bias (mA)	Subharmonic Frequency (GHz)	Subharmonic Power (dBm)	Fundamental Frequency (GHz)	Fundamental Power (dBm)
			Iris circuit		
1	115	28.13	12.05	55.02	9.5
3	140	29.05	12.0	57.84	9.9
5	140	29.51	10.2	59.5	7.3
6	150	34.48	16.9	69.1	11.5
7	150	32.25	15.2	63.93	11.1
8	150	32.98	14.7	66.25	10.6
9	150	31.91	11.1	63.7	7.6
			Resonant Cap Circuit		
10	150	33.4	14	67.0	10.5
10	160	36.0	14	72.2	11.1

guide as previously described, so that no dielectric-loading was necessary.

In both circuits, the subharmonic power was maximized by reactively terminating the fundamental power, and vice versa. The subharmonic power is generally greater than that at the fundamental output by about 3 to 6 dB. This is in general agreement with Johnston, Schafetter, and others.[19–20] However, the current density used here is the same as for the diodes operated in the fundamental frequency mode, and the frequencies are harmonically related to each other.

VI. CONCLUSIONS

By using an idler resonance at the subharmonic frequency, one can design oscillators with 20 percent tuning range. Such tunable oscillators and frequency deviators were built and worked satisfactorily in an experimental millimeter-wave PCM repeater system.

Multiple frequency circuits for IMPATT diode oscillators offer an important means of generating millimeter waves with useful power. For example, 2 mW of power at 3/2 the transit-time frequency, or 86 GHz, has been obtained.

VII. ACKNOWLEDGMENT

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